Radar having a transmission oscillator which can be excited

quasi-phase coherently by an evaluation oscillator

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Many forms of radar sensors are known for use in industry and

transport. Radar sensors are designed in the form of pulsed

radar, as described in US 3,117,317, US 4,132,991 and US

4,521,778, or in the form of so-called continuous wave or CW

10   radars. Pulsed radars determine the distance between an object

and the radar by measuring the delay time of a short pulse

from the radar to the object and back. In contrast, CW radars

evaluate the phase difference between the transmitted signal

and the reflected received signal. In order to determine a

15   reliable measurement value in a relatively wide measurement

range, the frequencies of CW radars are generally swept over

as wide a frequency range as possible. The best known and most

widely used radar of this type is the FMCW radar (FMCW =

frequency modulated continuous wave).

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Particularly in free-space applications, it is desirable to be

able to reduce the transmission power of a radar sensor

deliberately, since the radio licensing regulations frequently

require that the emitted transmission power and, in

25   particular, the interference transmissions associated with the

transmission process, remain below a specific level.

EP 1 051 639 B1 proposes a pulsed FMCW radar which is distinguished in that the radio-frequency output power of the radar can be influenced by the choice of a pulse sequence with

5   which the transmission signal is clocked.

Figure 1 illustrates a circuit variant according to the prior art as an improvement to the radar disclosed in EP 1 051 639 B1. The radar illustrated here has, like a conventional FMCW

10   radar, a voltage controlled oscillator VCO, a directional coupler RK, an antenna ANT and a mixer MIX1. In the same way as described in EP 1 051 639 B1, the radar also has clock logic CLK, by means of which the transmission signal is switched on and off periodically via a switch SWTX. For

15   example, if the switch is closed at a repetition rate of 1 MHz for only 1 ns in each case, then the mean transmission power is decreased by 60 dB in comparison to continuous operation. Any other desired values can also be set within a wide range by variation of the duty ratio.

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The circuit variant also has a second mixer MIX2. The signal received via the antenna ANT is keyed or amplitude-modulated by the clock logic CLK located in the transmission path, and has the spectrum illustrated (on the left) in Figure 6. This

25   modulation is precisely counteractive by mixing the received signal with $s_{CLK}(t)$ in MIX 2, and the measurement signal s1(t)

is converted to the signal s2(t) in baseband that is to say in the frequency range around the frequency 0.

A measurement signal as in the case of conventional FMCW radars which are operated without being pulsed is thus produced after filtering by means of the low-pass filter TPNF, which suppresses higher mixing products.

A concept as illustrated in Figure 1 has the advantage that the clock rate of the clock logic CLK can be varied over wide ranges. The width of the variation range is limited only by the bandwidth of BPZF and BPC. The magnitude of the mean transmission power can thus also be varied over a wide range without changing the form or frequency of the measurement signal. All that is changed is the signal-to-noise ratio of the measurement signal.

The embodiment shown in Figure 1 has the disadvantage that the switch SWTX must in this case be a radio-frequency switch.

On the one hand, radio-frequency switches are very complex, and on the other hand high switching contrast can be achieved only with difficulty. In consequence, the power can in practice often be reduced only to a restricted extent by means of the circuit shown in Figure 1.

A variant according to the prior art whose circuitry complexity is considerably simplified and which solves the problem is shown in Figure 2. In this case, the switch SWTX switches the oscillator itself on and off. This can be

5   achieved, for example, by connection and disconnection of the supply voltage. If the oscillator is not oscillating, it does not produce any signal power, and the switching contrast is a maximum. Furthermore, a so-called transmission mixer TRMIX is used instead of the directional coupler, in order to simplify

10   the circuit.

However, the circuit shown in Figure 2 has the disadvantage that either the maximum measurement range or the minimum mean transmitted power or the minimum measurement rate of the radar

15   is restricted by this type of switching. This is because, on the one hand, the switch SWTX must in this case always remain closed until all of the signal components, that is to say even those from the maximum distance to be expected or from the maximum predetermined distance, have been received again. The

20   pulse duration, that is to say the time during which the switch is closed and the oscillator is switched on should thus be considerably longer than the maximum signal delay time. With a measurement range of, for example, 15 m, the maximum signal delay time would be expected to be 70 ns, so that in

25   this case sensible measurement values would be expected only with pulse durations of more than 100 ns. In order on the

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other hand to achieve a desired reduction in the mean power by means of the duty ratio, correspondingly long pauses must be ensured between the switching-on moments, and in some cases these contradict the sampling theorems. This is because each

5   period for which the radar is switched on corresponds to the production of one sample value, which corresponds to the sampling of the measurement signal of a continuously operated FMCW radar. It is thus impossible to evaluate high frequencies if the pauses are too long. The distance to a measurement

10 object, to which the measurement signal frequency is proportional, must therefore not exceed a specific distance at a predetermined sweep rate of the frequency ramp. If the measurement range is nevertheless increased beyond this level, the sweep rate must be reduced, and the measurement time for a

15 measurement must therefore be increased.

One object of the present invention is to describe the systems which achieve the object of the described apparatuses in a different and improved form and whose transmission and

20 operating behavior can, furthermore, be controlled particularly advantageously.

This object is achieved by the invention specified in the independent claims. Advantageous refinements are specified in

25 the dependent claims.

An arrangement or apparatus accordingly has transmission means for production and transmission of an electromagnetic signal, and reception means for reception of a reflection of a transmitted electromagnetic signal. In this case, the

5 transmission means have a transmission oscillator for production of the electromagnetic signal, and the reception means have an evaluation oscillator for production of an electromagnetic valuation signal which can be compared, in particular mixed, with the received reflection of the

10 transmitted electromagnetic signal. The transmission oscillator and the evaluation oscillator are in this case connected and/or arranged such that the transmission oscillator can be excited quasi-phase coherently and/or is coupled quasi-phase coherently to the oscillating evaluation

15 oscillator by electromagnetic interaction.

In consequence, the transmission oscillator need be switched on only when required, thus reducing the energy consumption and the mean transmission power, as desired. On the other

20 hand, however, the quasi-phase coherent excitation or coupling always ensures sufficient phase coherence between the transmission oscillator and the evaluation oscillator, and the electromagnetic signal which is generated by the transmission oscillator, as well as between its reflection and the

25 evaluation signal which is generated by the evaluation oscillator.

Alternatively or additionally, the evaluation oscillator can also be excited quasi-phase coherently by the transmission oscillator for other purposes.

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In this case, quasi-phase coherently means that the arrangements according to the invention ensure that the phase difference between the signals from the transmission oscillator and the evaluation oscillator is small, but not

10 necessarily negligible, as in the case of pure coherence. The expression small phase difference should in this case be considered with respect to the intended communication or measurement task. By way of example, the value $\pi/10$, that is to say approximately 20°, is frequently used as the limit for

15 a small phase error. Signals such as these with only small phase errors are referred to in the following text as being quasi-phase coherent, and the time interval in which this coherence exists is referred to as the coherence time period.

20 One fundamental idea for implementation of quasi-phase coherence is described in DE 100 32 822 A1 and comprises an oscillator initially being in unstable equilibrium after being switched on, and first of all having to be caused to oscillate by some sort of external power supply. An oscillation starts,

25 via the feedback, only after this initial stimulus. This

initial stimulus is normally provided, for example, by thermal noise. This means that an oscillator starts to oscillate with a random phase and amplitude, at the frequency which is predetermined by its resonant circuit. However, if an external excitation signal is injected into the oscillator when it is switched on, then the oscillator does not start to oscillate randomly, but deterministically with respect to the phase of the injected signal. As long as the phase difference between the injected signal and the oscillator signal is less than about 20°, the two signals are quasi-phase coherent.

The characteristic of an oscillator attempting to follow the phase profile of a stimulating signal in the switching-on phase is a fundamental physical characteristic of any oscillator, which characteristic is, however, used according to the invention to couple two oscillators to one another quasi-coherently and to operate the transmission oscillator quasi-coherently with respect to the evaluation oscillator, or else, alternatively, to operate the evaluation oscillator quasi-coherently with respect to the transmission oscillator.

For the quasi-phase coherent excitation capability, the oscillators are electromagnetically coupled to one another in such a way that the transmission oscillator can be excited quasi-phase coherently by the evaluation oscillator, and/or the evaluation oscillator can be excited quasi-phase

coherently by the transmission oscillator. This can be achieved by the oscillators being connected to one another by line means. On the other hand, the crosstalk from one oscillator to the other is generally sufficient, provided that

5   the oscillators are arranged close to one another and, in particular, are not shielded from one another.

The oscillator which can be excited quasi-phase coherently is preferably repeatedly brought to the state in which it can be

10   excited quasi-phase coherently by being switched off and on again repeatedly at a clock rate by means for cyclic switching of the oscillator which can be excited quasi-phase coherently. In general, the pulse sequence should be chosen such that the time for which the oscillator which can be excited quasi-phase

15   coherently is switched on is shorter than or equal to the time period in which the quasi-phase coherence exists between the oscillator which can be excited quasi-phase coherently and the oscillator which provides the quasi-phase coherent excitation. The clock repetition rate at which the oscillator which can be

20   excited quasi-phase coherently is switched is, in particular, considerably higher than the maximum measurement signal frequency to be expected, in particular more than five times as high.

25   The time for which the oscillator which can be excited quasi-phase coherently is switched on is in this case in the same

order of magnitude as the reciprocal, or is less than the reciprocal, of the maximum difference to be expected between the frequencies of the two oscillators, which are coupled quasi-phase coherently, in the steady state.

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The frequencies of the transmission oscillator and/or of the evaluation oscillator are variable. The arrangements can then be operated as a frequency-modulated pulse radar or as a pulsed FMCW radar.

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One variant, which is very particularly suitable for short range, is obtained by the frequency of one oscillator being variable, while the other oscillator is in each case a fixed-frequency oscillator.

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The arrangement is, in particular, an arrangement for distance measurement and/or a radar, in particular a radar which can be swept through a frequency range and/or a radar which can be switched to two switching states between a short-range radar
20 and a long-range radar.

In a measurement method, in particular for distance measurement,
- a transmission oscillator is excited quasi-phase coherently
25 by an evaluation oscillator and/or an evaluation oscillator is excited quasi-phase coherently by a transmission oscillator,

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- a signal to be transmitted is produced by a transmission
oscillator,

- the signal is transmitted,

- a reflection of the transmitted signal is received, and

5 - an evaluation signal is produced by the evaluation
oscillator.

The reflection of the transmitted signal is then evaluated in
the measurement process, in particular taking into account the

10 evaluation signal. The evaluation signal is in this case
preferably mixed with the reflection of the transmitted
signal.

It is furthermore preferable for the transmission oscillator

15 to be excited quasi-phase coherently repeatedly and cyclically
by the evaluation oscillator, and/or the evaluation oscillator
is excited quasi-phase coherently, repeatedly and cyclically
by the transmission oscillator, for which purpose the
oscillator which is excited quasi-phase coherently is switched

20 off, and is switched such that it can be excited quasi-phase
coherently again, at a clock rate, for example by switching it
off and on.

Other advantageous refinements of the method result

25 analogously to the advantageous refinements of the
arrangement.

Further advantages and features of the invention are explained in the description of exemplary embodiments. In the figures:

5   Figure 1 shows a radar according to the prior art;

Figure 2 shows a second radar according to the prior art;

Figure 3 shows an arrangement with transmission means and
10   reception means;

Figure 4 shows an alternative arrangement with transmission means and reception means;

15   Figure 5 shows a. further arrangement with transmission means and reception means;

Figure 6 shows a spectra of a measurement signal at various stages in the measurement signal path;
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Figure 7 shows yet another arrangement with transmission means and reception means;

Figure 8 shows a demodulator;
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Figure 9 shows an alternative demodulator;

Figure 10 shows schematic measurement spectra of the arrangement as shown in Figure 7 and of a conventional rangefinding radar by comparison;

5   Figure 11 shows measurement results for the arrangement shown in Figure 7, and for a conventional rangefinding radar by comparison;

Figure 12 shows embodiments of a converter, and

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Figure 13 shows variants of the coupled signal sources.

The problems relating to Figure 2 are solved by the circuit shown in Figure 3. In addition to the oscillator VCO, which is

15   now used as the evaluation oscillator, the circuit contains a further oscillator, the transmission oscillator, LO, which is used to produce the transmission signal. The critical factor is that the evaluation oscillator VCO and the transmission oscillator LO are coupled to one another in a quasi-phase-

20   locked manner or quasi-phase coherently via the coupling QPK. If the transmission oscillator LO is switched on cyclically while the signal from the VCO is coupled to it, then the transmission oscillator LO repeatedly starts to oscillate with a phase which is predetermined by the evaluation oscillator

25   VCO.

A loose connecting line which is coupled to both oscillators may be provided in order to ensure this coupling QPK. However, because oscillator power is always being emitted, there is generally no need to provide specific circuitry apparatuses in

5   order to couple the oscillators to one another, provided that the oscillators are not shielded from one another, that is to say, for example, they are each designed without their own electrically closed metallic cover. Crosstalk from the evaluation oscillator VCO to the transmission oscillator LO

10   naturally likewise takes place via the ground or supply lines. No specific requirements need be observed in the choice of the oscillators, either. Conventional radio-frequency oscillators may be used, although one of the two oscillators must, according to the invention, be switched off and on cyclically

15   by a suitable pulse sequence in order in this way, as described, to ensure quasi-phase coherent coupling of the two oscillators. Care should preferably be taken to ensure that there is no third significantly stronger signal acting on the switched transmission oscillator LO from the outside and that

20   this oscillator does not start to oscillate coherently in response to this third signal instead of that from the evaluation oscillator VCO.

If the frequencies of the oscillators LO and VCO are the same,

25   then s3(t) becomes DC voltage which represents the difference phase (which by definition is small) between the two quasi-

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coherent output signals. If the transmission oscillator LO is in each case switched on cyclically for only a very short time, the quasi-phase coherence is maintained between the two oscillators over all of the periods in which they are switched

5   on. This means that the signal s3(t) of the mixer MIX2 is virtually identical to the signal which would result if the signal from the oscillator were keyed by a switch at the same clock rate.

10   As is illustrated in Figure 13, the evaluation oscillator VCO may also additionally be switched, in addition to the transmission oscillator LO. In this case, its stabilization must be largely complete at the moment at which the transmission oscillator LO is switched on, and it must not be

15   switched off again until the reflections have arrived from the maximum range area detected by the sensor. The switching of the evaluation oscillator VCO does not change the phase conditions between the transmitted signal and the received signal with respect to a measurement with the evaluation

20   oscillator VCO switched on all the time. However, this operating mode is advantageous for energy consumption reasons. In the arrangement shown in Figure 3, it is not relevant that the transmitting antenna ANTTXP and the receiving antenna ANTRX are formed separately. A single transmitting/receiving

25   antenna could likewise be fed via a directional coupler or circulator. The use of a PLL synthesizer or a PLL for

controlling the VCO (PLL = Phase locked loop) is not critical either. All that is indicated here is the capability to completely control the sensor by means of a microprocessor MP.

5   If the oscillator LO is a fixed-frequency oscillator and the VCO is a fixed-frequency oscillator, this results on the one hand in a constant frequency offset and a sin(x)/x amplitude weighting of the measurement signal s3(t). However, both effects can be taken into account in the measurement signal
10   evaluation process quite easily by calculation.

The circuit shown in Figure 4 produces the same measurement signals as the circuit shown in Figure 1. In this case, the variable-frequency transmission oscillator VCO2 is designed to
15   be as identical as possible to the variable-frequency evaluation oscillator VCO1, and the two control inputs of the oscillators are connected to one another. In the same way, as in Figure 3, the transmission oscillator VCO1 and the evaluation oscillator VCO2 are coupled to one another via
20   quasi-phase-locked or quasi-phase coherent coupling QPK.

If the transmission oscillator VCO1 and the evaluation oscillator VCO2 are of the same type, then the frequencies of the two oscillators VCO2 and VCO1 match virtually precisely
25   over the entire frequency sweep range, that is to say at any switching-on time. Since both oscillators now oscillate quasi-

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coherently at the same frequency over the entire frequency
sweep range, this results in a signal s4(t) as in the case of
a continuously operated FMCW radar, with only the signal-to-
noise ratio being reduced.

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In this case as well, it is once again possible as shown in
Figure 13 for the evaluation oscillator likewise to be
designed to be switchable so that the sensor consumes less
energy.

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A suitable configuration could appear as follows: the mid-
frequency of the oscillators could, for example, be in the
vicinity of 6 GHz or 26 GHz. By way of example, the oscillator
VCO1 is tuned in a bandwidth of about 1 GHz (or even more)

15   within 1-10 ms. The CLK clock rate can then, for example,
preferably be in the range from 1 to 10 MHz. The pulse
duration is then preferably in the range from 100 ps to 10 ns.

Furthermore, the operation of the sensor according to the

20   invention as shown in Figure 3 or 4 is maintained when, as
shown by dashed lines, a frequency converter UMSET with
embodiments as shown in Figure 12 is inserted in the sensor.
As illustrated on the left in Figure 12, the converter has a
fixed-frequency oscillator HFO, one mixer in each case and,

25   possibly, a bandpass filter in the transmission path and
reception path. The bandpass filter may be required in order

to suppress undesirable mixing products, for licensing

reasons. The fixed-frequency oscillator HFO may once again be

designed to be switchable for energy reasons, in which case

the oscillator must have largely stabilized when the

5   oscillators LO and VCO1 are switched on, and must not be

switched off until the reflections have arrived at the sensor

from the maximum range area to be detected. Alternatively, as

is illustrated on the right in Figure 12, the implementation

may have two separate oscillators HFO1 and HFO2 which are

10   coupled quasi-coherently to one another. In this case, the

switching-on time can once again be optimized for coherence

and energy consumption purposes.

A more advanced refinement of the circuit in the form of a

15   multifunction radar is shown in Figure 5. By way of example, a

switching signal SW-P/CW from the microprocessor MP can in

this case be used to switch the radar to two operating states.

In the switching state 1, the radar operates as a conventional

FMCW radar with full output power and, accordingly, a long

20   range and high sensitivity. In this operating mode, the

microprocessor control system is used to ensure that: a) the

switch SWTX is open, that is to say the oscillator VCO2 does

not produce a signal, and b) that the oscillator VCO1 produces

only signals in the frequency range which comply with the

25   licensing regulations, that is to say by way of example

according to FCC 15 in the ISM band from 24 GHz to 24.25 GHz

with an emitted field strength of < 250 mV/m.

In the switching state 2, the sensor operates as an ultra wideband pulsed FM radar in accordance with the extended licensing regulations in FCC part 15 "Ultra-Wideband Transmission System". Different antennas are also preferably used for the different operating modes. In the exemplary embodiment, the antenna ANTTXP has a considerably wider beam angle than the antenna ANTTXFM. This is worthwhile because, in the operating mode 2, this system is preferably suitable for use as a short-range radar while, in the switching state 1, it is preferably used to cover longer ranges. The antenna ANTRX may also be designed to be switchable between these two directionalities or, for the sake of simplicity, may be designed in the same way as the antenna with the larger beam angle (that is to say in the same way as ANTTXP). The antennas may be separate, joint or switchable, and array elements may be connected or disconnected in antenna arrays. Those skilled in the art will normally be familiar with the ways in which such antenna solutions and switches are implemented, and this is therefore not the subject matter of the statements here.

The radar as shown in Figure 5 is ideally suitable for use as a motor vehicle radar. It covers the short-range area in the operating mode 2, for example for use as a parking aid. In the operating mode 1, the radar sensor can carry out measurements

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at significantly longer distances, for example up to typically 100 m, admittedly with poorer resolution, and can thus be used well for the purpose of speed control and collision avoidance (for example as so-called Stop and Go Radar). The switching

5   between the operating modes could, for example, be coupled to the gear selection; reverse and first gear result in the operating mode 2, and all the other gears in the operating mode 1. Coupling to the vehicle speed or continuously varying operation would also likewise be feasible.

10

All the described radar are, of course, not limited to specific applications but may be used in all fields where pulsed or CW radars would also otherwise be used.

15   A method for operation of the sensor as shown in Figure 3 and the method for evaluation of its measurement signals will now be described in the following text. The signals from the other arrangements can be described as in the text above or can be evaluated in the conventional way for normal radar sensors.

20   However, in the case of the radar sensors shown in Figure 3, there are a number of special features which result in particularly advantageous capabilities.

Figure 7 once again shows the basic sign of the sensor and the

25   designations of the signals.

The function of the advantageous method for range measurement can be derived as follows: first of all, a single-frequency signal in the form

5    $s_{VCO}(t) = \sin((\omega_{sw}) \cdot t + \phi_0)$

is assumed as the signal from the oscillator VCO $s_{VCO}(t)$, where ω_c is the mid-frequency, ω_{sw} is an initially fixed modulation frequency, t is the time and ϕ_0 is an undefined phase offset.

10   As described above, the transmission oscillator LO is switched on and off cyclically. The frequency at which the oscillator LO is switched on and off is referred to in the following text as f_{mk}, and the time period for which the oscillator is in each case switched on is Ts.

15

During each switching-on process, the oscillator LO in the preferred arrangement starts to oscillate exactly with the instantaneous phase of $s_{VCO}(t)$, at its own natural oscillation frequency ω_{LO}. If, by way of example, the oscillator is

20   switched on at the time $t = -\tau$, then it starts to oscillate with the phase

$\phi_1 = \arg\{s_{VCO}(-\tau)\} = (\omega_c + \omega_{sw}) \cdot (-\tau) + \phi_0$

25   and, in consequence, the oscillator signal $s_{LO}(t)$ corresponds

to:

$$s_{LO}(t) = \sin(\omega_{LO} \cdot t - (\omega_c + \omega_{sw}) \cdot \tau + \phi_0).$$

5   This signal is transmitted from the radar to the respective

reflector, is reflected on the latter and arrives at the radar

delayed by the delay time τ, where τ = 2*dist/c, with dist

being the distance between the radar and the reflector and c

being the speed of light, as the received signal $s_{rx}(t)$ =

10 $s_{LO}(t-\tau)$.

$$s_{rx}(t) = s_{LO}(t-\tau) = \sin(\omega_{LO} \cdot (t-\tau) - (\omega_c + \omega_{sw}) \cdot \tau + \phi_0)$$

This received signal $s_{rx}(t)$ is mixed with the instantaneous VCO

15 signal $s_{VCO}(t)$ in the mixer. If the high-frequency mixing

products are ignored and, for simplicity, it is assumed that

$\omega_{LO} = \omega_c$ as is possible without any restrictions to generality

if ω_{sw} is chosen appropriately, then the mixed signal $s_{mix}(t)$

becomes:

20

$$s_{mix}(t) = \cos(t \cdot \omega_{sw} + \tau \cdot (\omega_c + \omega_{sw})).$$

The following text is now based on the assumption that

electronic components/means DEMOD are provided downstream from

25 the receiving mixer MIX1 and leads to the elimination of the

changeover time in the voltage in the time interval between switching on and switching off, that is to say from 0..Ts, in the sense of averaging. A simple envelope curve demodulator according to the prior art, in which the signal is detected

5   and is then low-pass filtered would, for example, operate in this sense.

Figure 8 shows one embodiment of a simple demodulator DEMOD such as this with a detector GR and the low-pass filter TP.

10

One more advantageous variant of a demodulator DEMOD, as is sketched in Figure 9, comprises the mixed signal preferably being down-mixed with a frequency in the vicinity of or (as is illustrated in Figure 1) equal to the cyclic frequency f_{mk} of

15   the clock control CLK/Sw to a low frequency, and then being filtered by means of a filter TP which has at least one low-pass response. One possible embodiment of this variant has a local oscillator LOZF, a mixer ZFMIX and a low-pass filter TP. In principle, the entire system now corresponds to the circuit

20   shown in Figure 3. If the frequency of the local oscillator LOZF is defined such that negative mixed frequencies can occur, then the mixer ZFMIX must be designed as a complex-value mixer, that is to say an IQ mixer according to the prior art (IQ: In Phase and Quadrature phase, that is to say 90°

25   phase-shifted). It is also possible, for example, to use a bandpass filter instead of the low-pass filter TP.

Considered in the time interval 0 to TS, the described means effectively result in the determination of the root mean square value of the voltage of the mixed signal $s_{mix}(t)$. This root mean square value from then on forms the actual measurement signal $s_{meas}(t)$. Constant amplitude factors are ignored in the following description. The root mean square value of $s_{mix}(t)$ in the time interval 0 to TS, that is to say $s_{meas}(t)$, is calculated as follows:

$$s_{meas}(t) = \int_0^{T_s} s_{mix}(t) = const. \cdot \frac{\cos\left(\omega_c \cdot \tau + \omega_{sw} \cdot \tau + \frac{1}{2} \cdot T_s \cdot \omega_{sw}\right) \cdot \sin\left(\frac{1}{2} T_s \cdot \omega_{sw}\right)}{\omega_{sw}}$$

Since the measurement system is preferably operated in a frequency-modulated form, the following text considers the situation in which the modulation frequency ω_{sw} is modulated as a function of time. If ω_{sw} is swept linearly from $-B/2$ to $+B/2$ over the bandwidth B during a time period T, that is to say:

$$\omega_{sw} = \frac{2 \cdot \pi \cdot B \cdot t}{T}$$

then, from $s_{meas}(t)$ for the resultant FMCW measurement signal $s_{measfmcw}(t)$:

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$$s_{meas_{fmcw}}(t) = \cos\left(\omega_c \cdot \tau + \frac{2 \cdot \pi \cdot B \cdot t \cdot \tau}{T} + \frac{\pi \cdot B \cdot t \cdot T_s}{T}\right) \cdot \frac{\sin\left(\frac{\pi \cdot B \cdot t \cdot T_s}{T}\right)}{\left(\frac{\pi \cdot B \cdot t}{T}\right)} \; .$$

As has already been described, the signal is reproduced

exactly only during a switched-on period in the derivation

5   with the signal $s_{measfmcw}$(t). Since this signal is additionally

modulated by the periodic modulation in the transponder, this

results in a frequency shift of $s_{measfmcw}$(t) by f_{mk} and

additional higher-order spectral components, as has already

been indicated in Figure 6. Since this amplitude modulation

10   effect is generally known, only a single spectral component

will be considered by way of example in the following text, or

the signal $s_{measfmcw}$(t), first of all as if it were not

cyclically modulated.

15   This measurement signal $s_{measfmcw}$(t) now has two critical and

highly advantageous differences in comparison to signals from

standard FMCW radars.

On the one hand, the measurement signal f_{meas}, which

20   corresponds to the derivative of the phase of the cos

argument, that is to say

$$f_{meas} = dist \cdot \frac{2 \cdot B}{T \cdot c} + \frac{B \cdot T_s}{2 \cdot T} = f_{beat} + \Delta b,$$

is shifted by the frequency component $\Delta b = B \cdot T_s / (2\ T)$. The frequency f_{beat} corresponds to the normal FMCW measurement frequency and includes the actual measurement information, specifically the distance between the radar and the reflector, where $\tau = 2\ dist/c$. On the other hand, the signal $s_{measfmcw}(t)$ is amplitude-weighted with a trigonometric function, in particular an Si function $(Si(x) = \sin(x)/x)$. The Fourier transform, that is to say the spectrum of this amplitude-weighted signal, results in a square-wave function, with the width Δp of the square wave being:

$$\Delta p = \frac{B \cdot T_s}{T}$$

Since the mid-frequency f_{meas} of the square wave occurs at $f_{beat} + \Delta b = f_{beat} + 0.5 \cdot \Delta p$ as a result of the described frequency shift, this results in the advantageous characteristic that that spectral component of the cos function which is located in the positive frequency range and in the negative frequency range does not overlap even when the range becomes negligibly small. This means that a radar system according to this invention can be used to measure ranges down to 0, in principle independently of the bandwidth. In consequence, this system does not have the short-range problem of comparable known radars with real-evaluation, whose basic spectrum is

illustrated in Figure 10.

This advantageous characteristic is assisted by the fact that
the spectral envelope curve is a square-wave function, as

5   illustrated at the top in Figure 10. In consequence, the beat
frequency f_{beat} and thus the distance can be determined on the
basis of the interval between the outer edges of the right-
hand and left-hand sidebands.

10   Figure 10 is intended for general illustrative purposes and
represents the frequency spectrum for the echo signals which
are received by the radar as measurement signals (echo
signals). The envelope curve of all the amplitudes $\hat{a}$ of the
echo signals over the entire frequency range or else over only

15   a section of it is referred to as the echo profile, which has
to be evaluated. Amplitudes $\hat{a}$, phases ϕ and frequencies f of
the echo signals are thus measured and are processed further
in a computation unit.

20   The system is preferably configured as follows: the switched-
on duration Ts is preferably chosen such that it corresponds
to half the reciprocal of the modulation bandwidth B, that is
to say Ts = 2/B. The modulation frequency f_{mk} is chosen on the
basis of the desired effective transmission path. One suitable

25   practical configuration could appear as follows:

$B = 1$ GHz, $Ts = 2$ ns and $f_{mk} = 2$ MHz.

It is particularly advantageous for the system according to

5   the invention to be used as a short-range radar. In the case

of a short-range radar, the described capability to use the

outer edge of the square-wave function to determine the beat

frequency or the distance is also particularly advantageous

since, in the case of sensors such as these, the shortest

10   distance or the closest object is frequently of particular

interest, for example when the sensor is used for collision

avoidance purposes. The measurement accuracy problems which

normally occur with systems according to the prior art and

result from echo superimpositions ideally do not occur with

15   the system according to the invention if, as proposed, the

inner edges of the square-wave function are used for

evaluation. The problems resulting from superimposition of

spectral components in conventional systems and the advantage

of the system according to the invention are illustrated

20   clearly in Figure 11. This shows the superimposition of 6

signal components (1, 2, ..., 6) illustrating that, in the

normal situation (at the top in Figure 11), the specific

measurement value, that is to say the maximum of the spectrum,

does not correspond with the actual frequency which is

25   proportional to the distance (in this case = 100).

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In contrast, in the case of the system according to the invention (at the bottom in Figure 11), the left-hand edge is located at the correct position even when the signal components are superimposed, that is to say a correct range
5   value is determined.

In order to exactly determine the edges, in particular the first innermost edge, it is advantageous to differentiate the magnitude spectrum, since the steep flanks of the square wave
10   then result in pronounced peaks. The positions of the maxima of the peaks then correspond to the position of the edges or points of inflexion of the edge function. This form of evaluation makes it possible to use simple means to avoid the amplitude of the signal influencing the location of the
15   specific position, as would be the case with simple threshold value evaluations. Since the measurement signals and spectra generally occur only in a time-discrete form, it is also advantageous to further improve the determination of the position of the maxima of the differential spectrum with the
20   aid of an interpolation calculation. An advantageous interpolation process can be carried out, for example, using a conventional polynominal fit. Some of the discrete spectral points in the immediate vicinity of the maximum (for example 3 points, that is to say the maximum and its right-hand and
25   left-hand neighbors) and an even-order polynominal (for example 2nd order, that is to say a parabel) are preferably

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used for this purpose. It is, of course, likewise possible to use other interpolators, such as spline methods or those methods which match the curve shapes to be expected to the measured curve by means of least-squares methods.

5

All of the methods mentioned may, of course, also be used individually, or the preferred features according to the invention may be developed in other combinations and may also be combined with other methods which are known for FMCW

10   radars.